AN-731
Application Note

LOW-SPEED MODEM FUNDAMENTALS

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This application note describes the interface circuitry and system performance of the MC6860 low-speed Modem. The basic design criteria for the bandpass filter, limiter, threshold detector, and duplexer are given. Evaluation of the system in the presence of gaussian noise is presented. This article also includes a brief description on data couplers.

MOTOROLA Semiconductor Products Inc.

LOW-SPEED MODEM FUNDAMENTALS

GENERAL

The MC6860 low-speed Modem can be used in many different configurations. These include full duplex, half duplex, simplex, automatic answering, automatic disconnect, originate only, answer only, answer/originate, and others. Figure 1 illustrates the basic modem configuration used to evaluate the MC6860. An originate only and an answer only modem design is used for evaluation, and each section of the interface circuitry is dealt with in this article.

The originate modem transmits on the low-frequency channel (Mark 1270 Hz and Space 1070 Hz) and receives on the high-frequency channel (Mark 2225 Hz and Space 2025 Hz). The answer modem transmits on the upper channel and receives on the lower.

A buffer and duplexer as shown in Figure 1 provide the modem interface to the transmission network while the bandpass filter allows only the desired receive signals to be seen by the limiter and demodulator.

MODULATOR – BUFFER

Mark/Space information that is presented to the Transmit Data input of the modem is converted to an FSK signal for transmission. The modulator output is an approximated sinewave derived from a digital-to-analog converter within the MC6860. There are eight amplitude levels per cycle. Each step has been optimized such that the composite waveform has a maximum amount of signal energy at the fundamental. Figure 2 shows the 1270 Hz transmit carrier and Figure 3 gives its spectral distribution. A nominal signal has the second harmonic attenuated to -30 dB.

The modulator output impedance is typically 2 k ohms. Loading this output with an impedance less than 100 k ohms can produce harmonic distortion. Therefore, a buffer amplifier is required to match impedances to the duplexer and the telephone line. This buffer amplifier may be designed to also provide filtering if additional clean-up of the transmitted signal is required.

The modulation spectrum for 300 bits per second using an alternate Mark/Space data format is shown in Figure 4. The amount of modulation or sideband energy that falls in the adjacent channel is an item of concern in full duplex operation. Under this condition both channels are operating simultaneously and all the adjacent channel energy that is not balanced out in the duplexer feeds directly through the bandpass filter and to the limiter. Excessive phase jitter results if the received signal level is low enough to approach that of the interference level at the limiter input. For this reason, additional filtering of the modulator output may be required before it feeds to the duplexer on those modem designs desiring wide dynamic ranges of input signal levels.

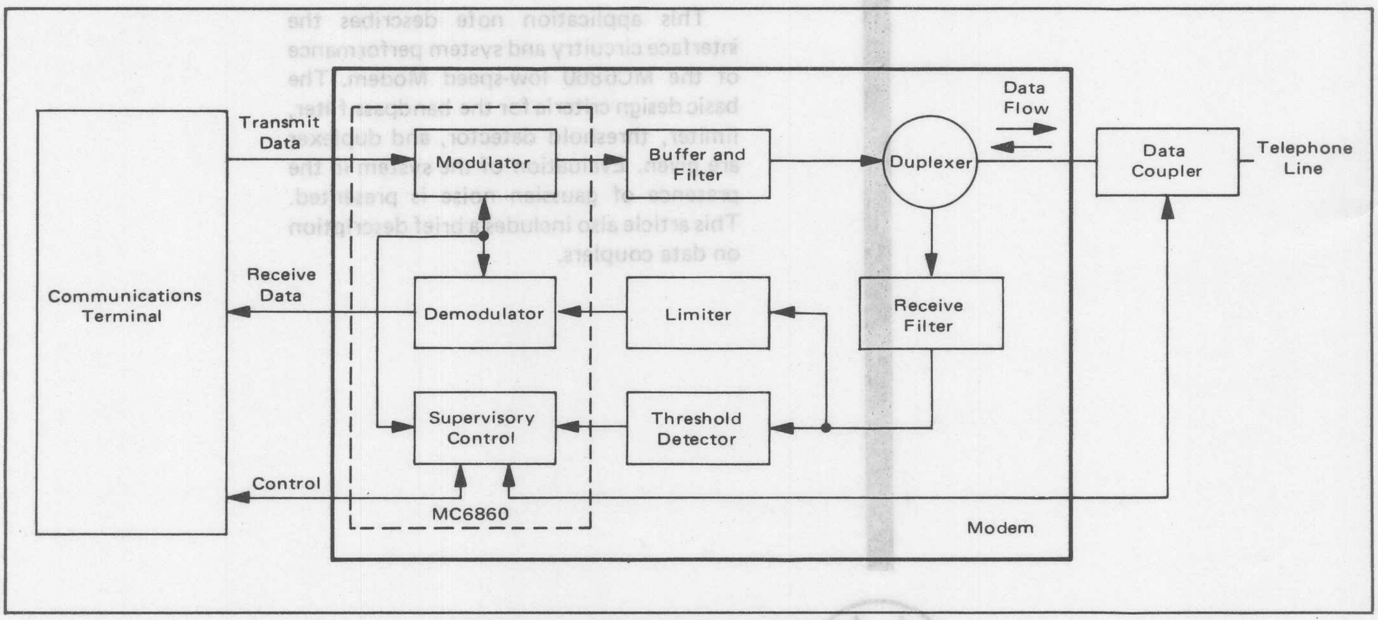

FIGURE 1 – Low-Speed Modem and Interconnections

Circuit diagrams external to Motorola products are included as a means of illustrating typical semiconductor applications; consequently, complete information sufficient for construction purposes is not necessarily given. The information in this Application Note has been carefully checked and is believed to be entirely reliable. However, no responsibility is assumed for inaccuracies. Furthermore, such information does not convey to the purchaser of the semiconductor devices described any license under the patent rights of Motorola Inc. or others.

Interference by the second harmonic is of concern in the originate mode only. In this mode, the transmit signal is in the low band and its second harmonic falls in or near the passband of the return channel. In half duplex operation, the transmit carrier is held at a constant Mark (1270 Hz) while data is being received. The second harmonic (2540 Hz), which is typically -30 dB or more below the fundamental in amplitude, falls just outside the passband of the receive filter and is further attenuated. In full duplex operation, the second harmonic and the modulation sidebands have about the same amount of energy. If this undesired energy must be reduced, the filter used to reduce the modulation sidebands will also reduce the second harmonic. Phase jitter and bias distortion inherent in the modulator is less than 3 μs.

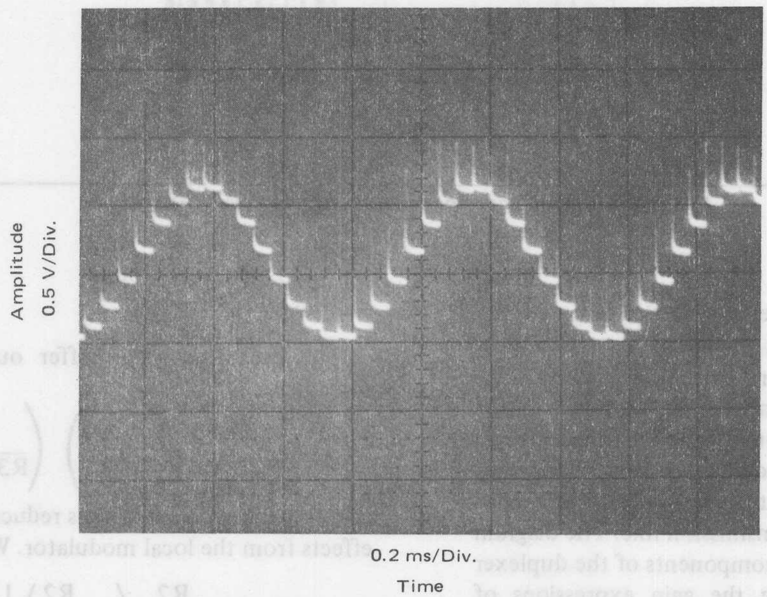

FIGURE 2 – MOS Synthesized 1270-Hz Sine Wave

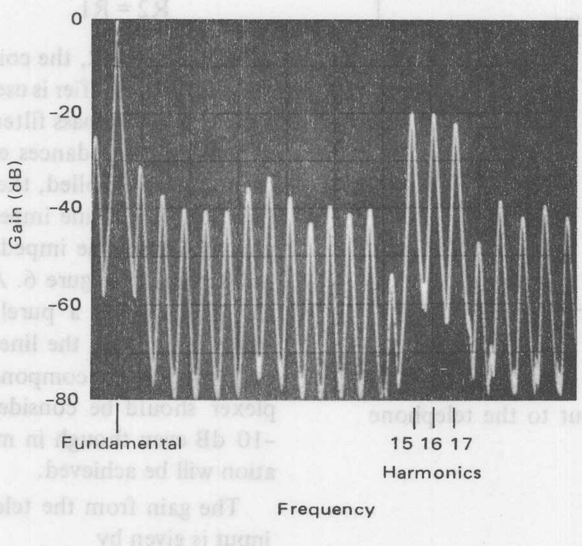

FIGURE 3 – Frequency Spectrum of MOS Sine Wave

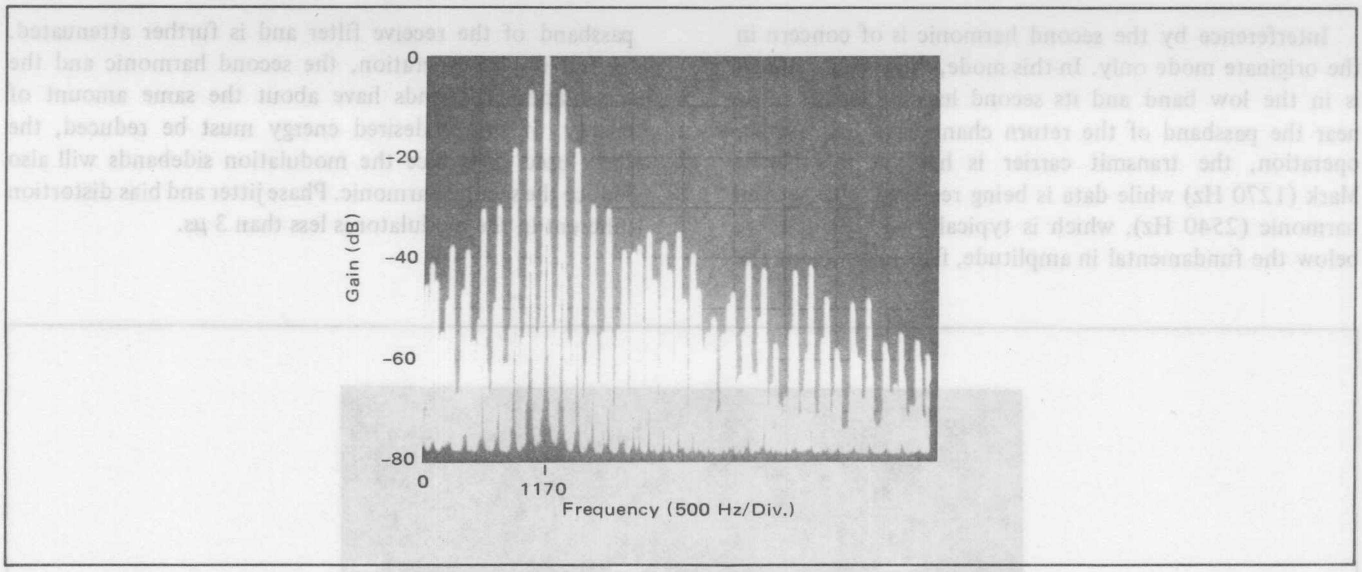

FIGURE 4 – Modulation Spectrum for Alternate Mark/Space

DUPLEXER

The duplexer is used to interface the modem with the transmission media which is a telephone system in most cases, through a data coupler. Since signal flow is bi-directional on the telephone line, the duplexer must allow the received signal to pass on to the bandpass filters, properly couple the transmitted signal onto the line, minimize the local transmit level at the bandpass filter input, and properly terminate the transmission line. The diagram of Figure 5 shows the various components of the duplexer with A1, A2, and A3 being the gain expressions of importance.

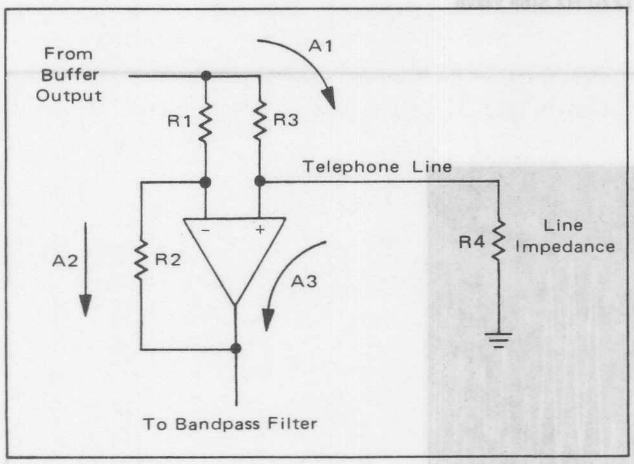

FIGURE 5 – Duplexer

The gain from the modulator output to the telephone line is

$$A1 = \frac{R4}{R3 + R4}$$

where R4 is the line impedance and is considered to be nominally 600 ohms resistive. Since the line must be properly terminated, R3 must equal R4. Therefore:

$$R3 = R4 = 600 \text{ ohms}$$

and $A1 = 0.5$

The gain from the buffer output to the bandpass filter input is

$$A2 = -\frac{R2}{R1} + \left(1 + \frac{R2}{R1}\right)\left(\frac{R4}{R3 + R4}\right)$$

It is desired that A2 = 0, thus reducing the intermodulation effects from the local modulator. With R3 = R4:

$$A2 = 0 = -\frac{R2}{R1} + \left(1 + \frac{R2}{R1}\right)\frac{1}{2}$$

$$2\frac{R2}{R1} = 1 + \frac{R2}{R1}$$

$$2R2 = R1 + R2$$

$$R2 = R1$$

With R1 = R2, the common mode characteristic of the operational amplifier is used to balance out the local modulator at the bandpass filter input, i.e., A2 = 0.

Since all impedances except the line impedance can be accurately controlled, the degree of nulling A2 becomes a function of the line impedance. The duplexer gain, A2, is plotted versus line impedance variation from 200 ohms to 1000 ohms in Figure 6. A well-defined notch exists when the line appears a purely resistive 600 ohms (the ideal case). In practice the line impedance can have reactive as well as resistive component variation, therefore the duplexer should be considered as providing approximately -10 dB even though in many connections greater attenuation will be achieved.

The gain from the telephone line to the bandpass filter input is given by

$$A3 = 1 + \frac{R2}{R1} = 2$$

when R1 = R2.

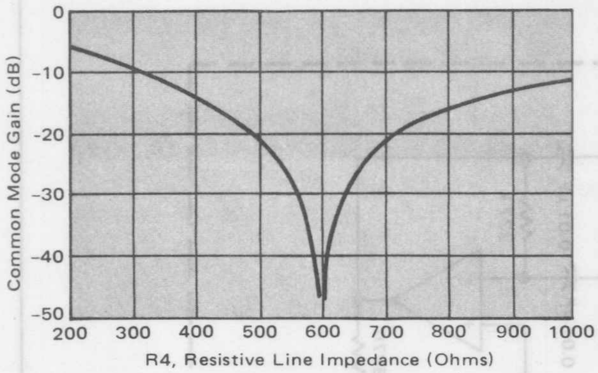

FIGURE 6 – Common Mode Gain versus Line Impedance

BANDPASS FILTER

The purpose of the bandpass filter is to amplify the received signal from the remote modem while rejecting all other signals that may be present in the local modem or on the telephone line. Interference which must be filtered out has several possible sources. Each of these must be considered and dealt with individually. Noise which is coupled in through the transmission media is either impulsive or band limited (gaussian) white noise. Both of these must be analyzed on a statistical basis. Discrete interfering signals may also be coupled in through the transmission media. However, the interfering signal of prime importance comes from the local modulator and will always exist in the half or full duplex modes.

Since the transmission media is lossy, the local transmit carrier level will exceed the level of the received signal. For this reason, the bandpass filter must have enough selectivity to reject the local carrier to an acceptable level. Modems that are designed for a wide dynamic range of input signal levels (-15 dBm to -55 dBm) require better than 70 dB rejection of interfering signals. Most of this rejection must come from the selectivity in the bandpass filter.

Reducing the effects of band limited white noise is accomplished by decreasing the bandwidth of the filter. Determining the minimum bandwidth comes by investigating the received signal characteristics. The transmitted data can be recovered from binary FSK by properly detecting the carrier and the first sidebands (first Bessel function)[1]. With a data rate of 300 bits per second and a data format of alternate Marks and Spaces, the first Bessel function occurs at ±150 Hz from the carrier. All other data formats have sidebands within the ±150 Hz limit. A minimum bandwidth of 300 Hz is then required in the bandpass filter.

The bandpass filter output is fed into an amplitude limiter, therefore the amount of passband ripple is not a critical parameter. An item of serious concern, however, is the phase linearity over the passband. All frequency components that pass through the filter must be equally delayed in time or jumbling and smearing of the data occurs. This is known as intersymbol or interbit interference. Performance of the communication system is degraded under these conditions with bias distortion and excessive phase jitter at the demodulator output resulting. Intersymbol interference can be reduced by linearizing the phase versus frequency transfer function. The slope of this transfer function is termed envelope delay and is determined by:

$$T_d = \frac{\Delta\phi}{\Delta f} \frac{1}{360 \text{ deg/cycle}}$$

where $\Delta\phi$ = change of phase in degrees
Δf = change of frequency in Hz

Minimizing the distortion of the envelope delay curve then minimizes the intersymbol interference. This is relatively easy over the center 2/3 of the passband. However, keeping constant delay near the band edges is quite difficult, if not impossible. For this reason, the optimum bandwidth is not determined according to the data rate but rather according to achievable linear phase characteristics. Bias distortion of one tenth of the bit period at 300 bps typically requires a -3 dB bandwidth of 450 Hz to 500 Hz.

Bandpass filters for evaluating the MC6860 were designed to have approximately a 450 Hz, -3 dB bandwidth with a Chebyschev response. The schematic for the answer filter is found in Figure 7 and is outlined for identification. The analytical response of this filter using standard valued components is tabulated in Table 1. The -3 dB bandwidth is calculated as 486 Hz and measured as 448 Hz. There is approximately 0.7 dB ripple over the center 300 Hz of the passband, with 0.4 ms envelope delay distortion, as shown in Figure 8. This filter attenuates the local transmit carrier of 2225 Hz by -35 dB relative to the passband gain.

A similar schematic for the originate bandpass filter is given in Figure 9. Its response approximates that of the originate filter as seen in Table 2 and Figure 10. Attenuation of the 1270 Hz local transmit carrier is -43 dB relative to the passband gain.

The envelope delay distortion for both of these filters can be reduced by widening the passband, thus flattening the envelope delay curve.

LIMITER-THRESHOLD DETECTOR

The demodulator in the MC6860 requires symmetrical limiting of the received signal in order to produce equal half-cycle periods. Each half-cycle period is measured in reference to an accurate time base to determine if the received frequency is a Mark or a Space. Non-symmetrical limiting produces errors in the demodulation process, thus degrading the system performance. Accurate limiting must be achievable over the expected input dynamic range. Such items as maximum input level and input offset voltage of the limiting device must be carefully considered.

Figure 11 shows the schematic for the limiter. The effect of the input offset is reduced by placing equal terminating resistors on both the inverting and noninverting outputs. An input coupling capacitor is used to block any dc bias coming from the output of the last amplifier of the bandpass filter. The desired ac signal is now properly centered about the input bias level of the

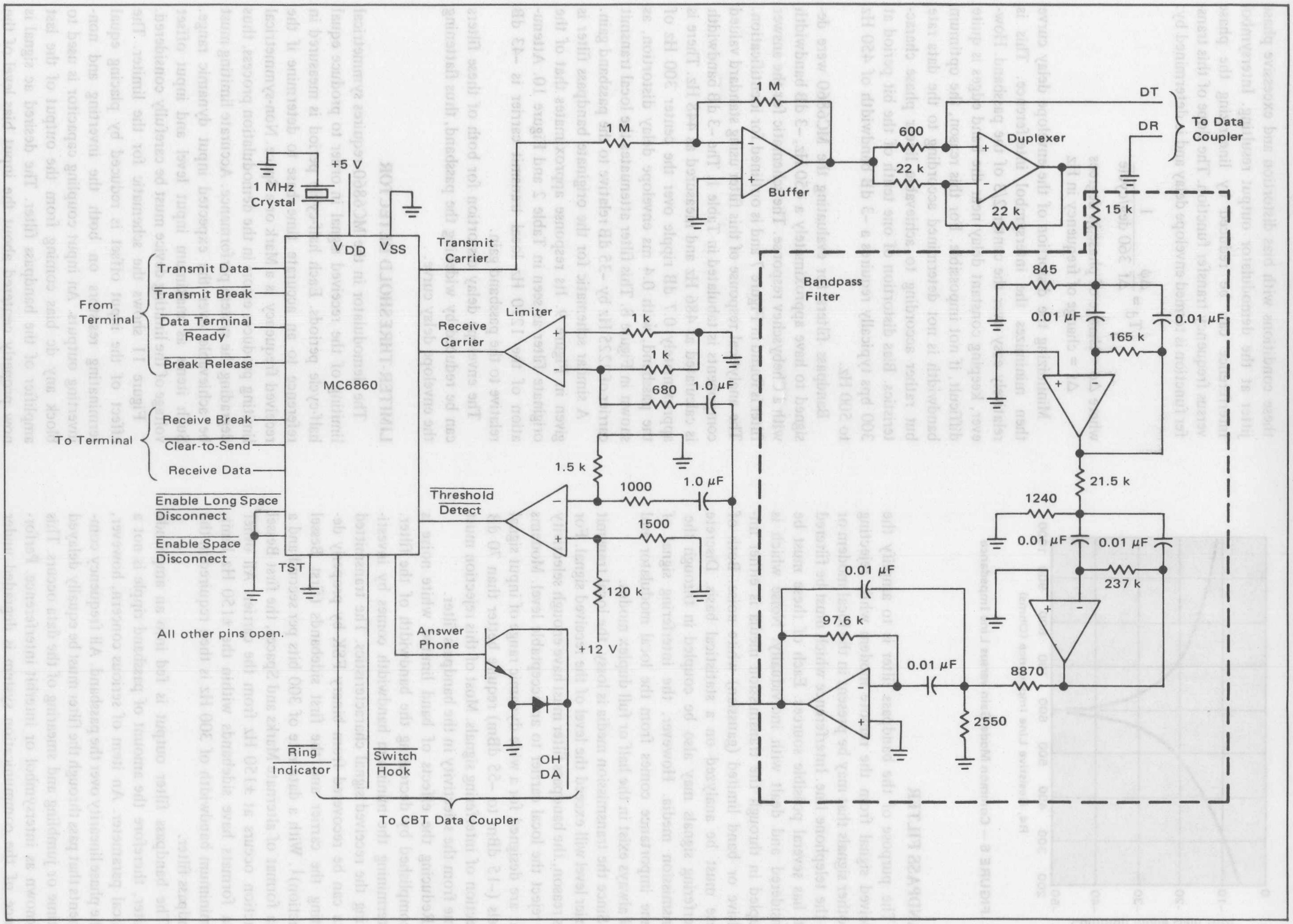

FIGURE 7 – Answer Modem

limiter and the maximum input dynamic range can now be achieved. A 40 dB dynamic range can be achieved with the limiter of Figure 11. Caution must be exercised in the amount of loading placed upon the bandpass filter output for distortion can result with large signal levels. An isolation resistor placed in series with the limiter input decreases the loading on the bandpass filter. Under maximum signal level conditions the limiter should be operating close to its upper input limit.

The output of the limiter is fed into the demodulator.

The threshold detector is used to determine if the input signal to the limiter is above the maximum detectable signal level of the modem. This is an amplitude measurement only, thus the period of the output is not critical. A comparator is used with one side biased to the peak amplitude of the desired minimum detectable signal level at the bandpass filter output. When the signal level exceeds the bias point, the comparator output goes low indicating an acceptable signal level.

TABLE 1 – Answer Filter Tabulated Response

Frequency	Node Voltage	dB Voltage	Phase Shift	Envelope Delay (ms)
0.3000E+03	0.202E–01	–33.883	80.83	
0.4000E+03	0.576E–01	–24.796	76.99	.11
0.5000E+03	0.145E+00	–16.770	72.24	.13
0.6000E+03	0.354E+00	–9.008	65.93	.18
0.7000E+03	0.906E+00	–.856	56.62	.26
0.8000E+03	0.267E+01	8.523	40.26	.45
0.9000E+03	0.101E+02	20.158	–1.03	1.15
0.9250E+03	0.141E+02	22.968	–21.78	2.31
0.9500E+03	0.176E+02	24.927	–47.11	2.81
0.9750E+03	0.194E+02	25.746	–72.41	2.81
0.1000E+04	0.196E+02	25.847	–93.83	2.38
0.1025E+04	0.194E+02	25.761	–111.32	1.94
0.1050E+04	0.193E+02	25.728	–126.37	1.67
0.1075E+04	0.195E+02	25.790	–140.24	1.54
0.1100E+04	0.198E+02	25.914	–153.70	1.50
0.1125E+04	0.201E+02	26.045	–167.08	1.49
0.1150E+04	0.203E+02	26.142	179.59	1.48
0.1175E+04	0.204E+02	26.190	166.42	1.46
0.1200E+04	0.204E+02	26.203	153.50	1.44
0.1225E+04	0.204E+02	26.210	140.79	1.41
0.1250E+04	0.205E+02	26.246	128.07	1.41
0.1275E+04	0.207E+02	26.321	114.90	1.46
0.1300E+04	0.209E+02	26.410	100.71	1.58
0.1325E+04	0.209E+02	26.417	84.93	1.75
0.1350E+04	0.203E+02	26.166	67.39	1.95
0.1375E+04	0.187E+02	25.459	48.87	2.06
0.1400E+04	0.163E+02	24.219	31.05	1.98
0.1425E+04	0.134E+02	22.568	15.50	1.73
0.1450E+04	0.108E+02	20.713	2.78	1.41
0.1500E+04	0.707E+01	16.984	–15.46	1.01
0.1600E+04	0.340E+01	10.635	–35.62	.56
0.1700E+04	0.193E+01	5.693	–46.41	.30
0.1800E+04	0.121E+01	1.703	–53.24	.19
0.1900E+04	0.829E+00	–1.634	–58.02	.13
0.2000E+04	0.596E+00	–4.501	–61.59	.10
0.2100E+04	0.446E+00	–7.016	–64.36	.08
0.2200E+04	0.345E+00	–9.255	–66.60	.06
0.2300E+04	0.273E+00	–11.275	–68.45	.05
0.2400E+04	0.221E+00	–13.115	–70.00	.04
0.2500E+04	0.182E+00	–14.807	–71.34	.04
0.2600E+04	0.152E+00	–16.372	–72.49	.03
0.2700E+04	0.128E+00	–17.830	–73.51	.03
0.2800E+04	0.109E+00	–19.194	–74.41	.03
0.2900E+04	0.947E–01	–20.476	–75.21	.02
0.3000E+04	0.824E–01	–21.687	–75.94	.02

TABLE 2 – Originate Filter Tabulated Response

Frequency	Node Voltage	dB Voltage	Phase Shift	Envelope Delay (ms)
0.3000E+03	0.467E–03	–66.607	87.38	
0.4000E+03	0.116E–02	–58.686	86.45	.03
0.5000E+03	0.242E–02	–52.315	85.47	.03
0.6000E+03	0.454E–02	–46.867	84.42	.03
0.7000E+03	0.794E–02	–42.001	83.28	.03
0.8000E+03	0.133E–01	–37.505	82.01	.04
0.9000E+03	0.218E–01	–33.233	80.59	.04
0.1000E+04	0.352E–01	–29.071	78.97	.05
0.1100E+04	0.567E–01	–24.925	77.07	.05
0.1200E+04	0.923E–01	–20.700	74.79	.06
0.1300E+04	0.153E+00	–16.298	71.98	.08
0.1400E+04	0.263E+00	–11.595	68.37	.10
0.1500E+04	0.477E+00	–6.425	63.50	.14
0.1600E+04	0.941E+00	–.531	56.43	.20
0.1700E+04	0.212E+01	6.534	44.84	.32
0.1800E+04	0.601E+01	15.574	20.75	.67
0.1825E+04	0.814E+01	18.210	9.86	1.21
0.1850E+04	0.110E+02	20.874	–4.80	1.63
0.1875E+04	0.146E+02	23.268	–24.14	2.15
0.1900E+04	0.176E+02	24.935	–47.21	2.56
0.1925E+04	0.192E+02	25.654	–70.39	2.57
0.1950E+04	0.194E+02	25.738	–90.48	2.23
0.1975E+04	0.191E+02	25.618	–107.18	1.86
0.2000E+04	0.189E+02	25.529	–121.58	1.60
0.2025E+04	0.189E+02	25.532	–134.82	1.47
0.2050E+04	0.191E+02	25.609	–147.63	1.42
0.2075E+04	0.193E+02	25.713	–160.44	1.42
0.2100E+04	0.195E+02	25.795	–173.34	1.43
0.2125E+04	0.196E+02	25.824	173.77	1.43
0.2150E+03	0.195E+02	25.798	161.04	1.41
0.2175E+04	0.194E+02	25.743	148.60	1.38
0.2200E+04	0.193E+02	25.696	136.37	1.36
0.2225E+04	0.193E+02	25.696	124.09	1.36
0.2250E+04	0.194E+02	25.756	111.29	1.42
0.2275E+04	0.196E+02	25.851	97.31	1.55
0.2300E+04	0.197E+02	25.876	81.50	1.76
0.2325E+04	0.191E+02	25.634	63.59	1.99
0.2350E+04	0.176E+02	24.888	44.47	2.13
0.2375E+04	0.150E+02	23.549	26.10	2.04
0.2400E+04	0.122E+02	21.768	10.27	1.76
0.2425E+04	0.976E+01	19.786	–2.47	1.42
0.2450E+04	0.775E+01	17.786	–12.48	1.11
0.2475E+04	0.621E+01	15.858	–20.37	.88
0.2500E+04	0.503E+01	14.039	–26.70	.70
0.2600E+04	0.247E+01	7.860	–42.90	.45
0.2700E+04	0.142E+01	3.031	–51.95	.25
0.2800E+04	0.901E+00	–.904	–57.83	.16
0.2900E+04	0.615E+00	–4.218	–62.00	.12
0.3000E+04	0.443E+00	–7.079	–65.14	.09

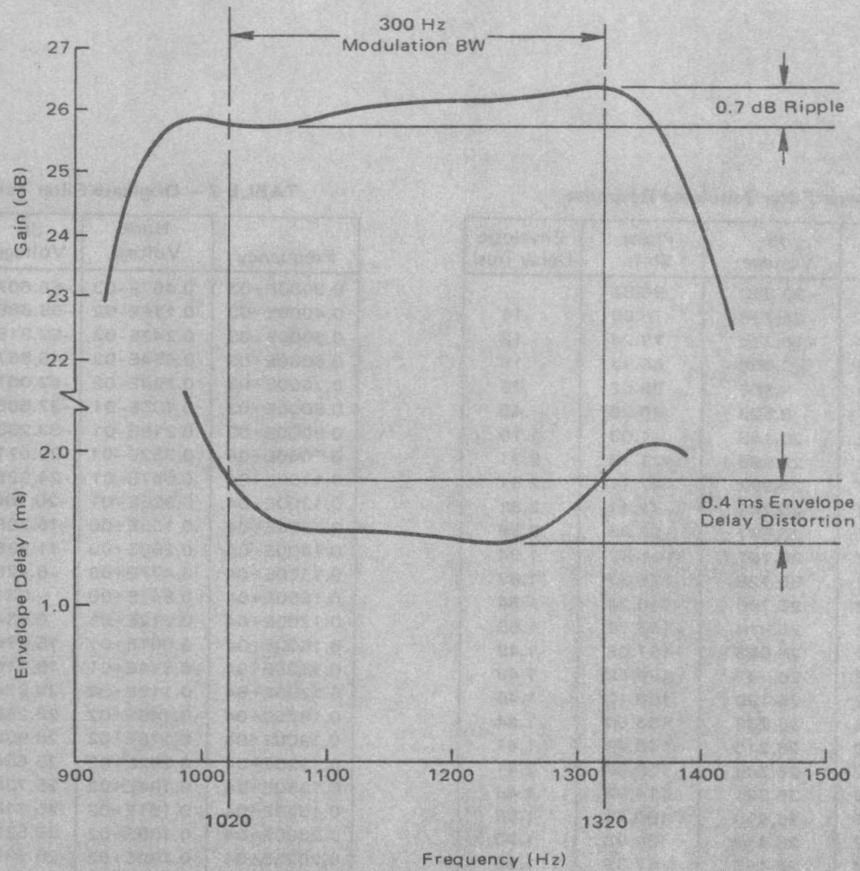

FIGURE 8 – Answer Bandpass Filter Characteristics

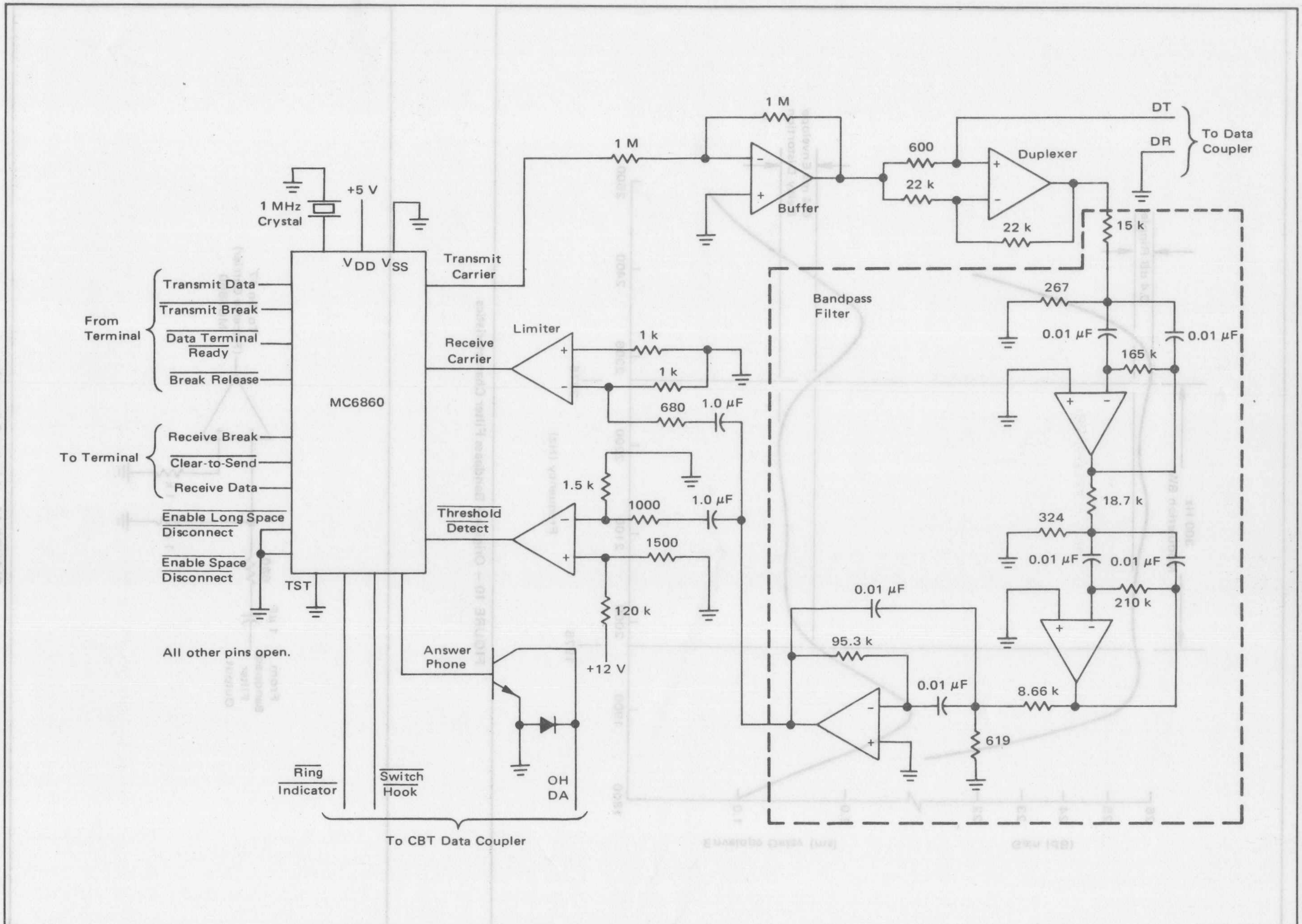

FIGURE 9 – Originate Modem

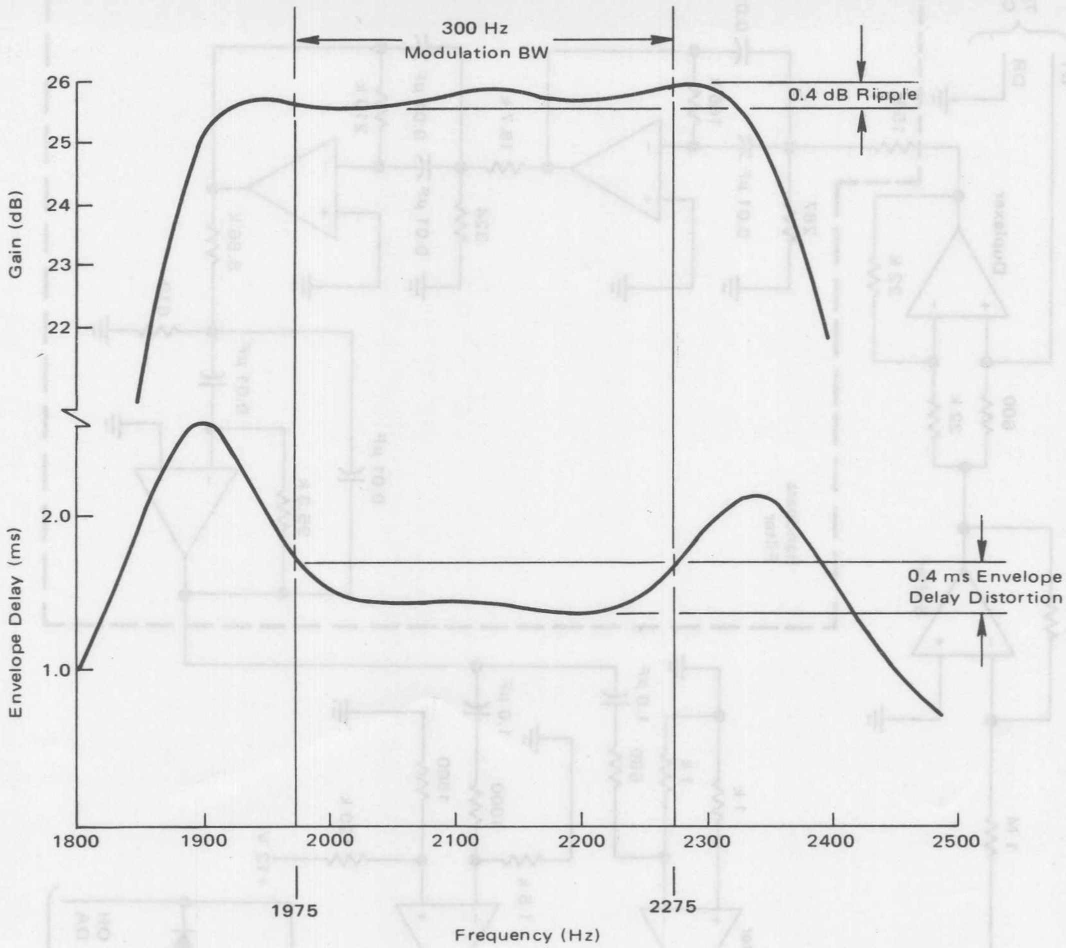

FIGURE 10 – Originate Bandpass Filter Characteristics

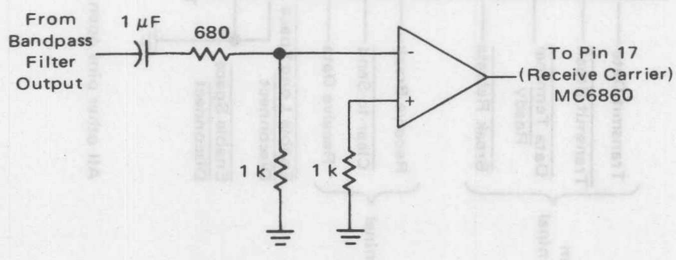

FIGURE 11 – Limiter Schematic

DEMODULATOR

The demodulator utilizes half-cycle detection for determining the presence of Mark or Space frequencies. Therefore, the Mark/Space information is quantized to half-cycle increments of the received carrier. Digitizing a linear signal produces a quantization error. This error appears in the form of phase jitter and bias distortion at the demodulator output of the MC6860.

The phase jitter of the demodulator output is shown in Figure 12. The upper trace is the alternate Mark/Space transmit data into the originate modulator. The lower trace shows the recovered data out of the demodulator of the answer modem. The inherent phase jitter of the demodulation process is approximated by

$$\%\phi \text{jpeak} \approx \frac{\text{Data Rate}}{4 \text{ Space Frequency}} \times 100$$

The receive Space frequency for the answer modem is 1070 Hz and the data rate is 300 bps, giving a peak phase jitter of 7%. This corresponds to 0.233 ms, as shown in Figure 12. The output Mark/Space transition will occur within 0.233 ms of the actual data transitions, neglecting bias distortion.

The receive Space frequency for the originate modem is 2025 Hz. The peak phase jitter is 3.7% (0.123 ms) at a data rate of 300 bps.

Bias distortion inherent in the demodulation process can be found according to:

$$\% \text{ Bias Distortion} \approx \frac{1}{2T}\left(\frac{1}{f_s} - \frac{1}{f_m}\right) 100$$

where T = Data bit period in seconds
f_s = Space frequency in Hz
f_m = Mark frequency in Hz

Thus the originate modem has a bias distortion of 0.67% and the answer modem has 2.2%. This is a marking bias (period of a Mark greater than period of a Space) for both modems.

Total distortion equals percent peak jitter plus percent bias distortion.

Careful inspection of Figure 12 reveals less than 0.2 ms marking bias. This is the accumulative bias distortion from the modulator input through the system to the demodulator output. The majority of this distortion results from the non-linear envelope delay through the bandpass filter in the answer modem. It is for this reason that special consideration must be given to delay distortion.

DATA COUPLERS

The two data couplers commonly used with low-speed modems are the CBS and CBT[2]. Each contains a data access arrangement (DAA) and the necessary telephone network control signaling functions. Figures 13 and 14 show the block diagrams of the CBS and CBT data couplers respectively. The supervisory control signals from the CBS comply with the RS-232 interface specifications, whereas the CBT control signals are contact closures and relay drive currents.

Table 3 identifies the various data coupler input/output signals.

SYSTEM PERFORMANCE

The MC6860 was evaluated in a typical system configuration. The tests utilized an originate only and an answer only design as outlined in Figure 15. The relative gains for both the answer and originate modems are given. A 600-ohm termination was provided to simulate the characteristic impedance of the transmission line, and to provide an input for the gaussian noise generator.

The test equipment was connected according to Figure 16. A word generator producing a 255-bit pseudo-random pattern at 300 bits per second was used as a transmit data input to the originate modem. The return channel was held at a constant Mark condition. The received data from the answer modem was compared for errors on a bit-by-bit basis with the transmitted data.

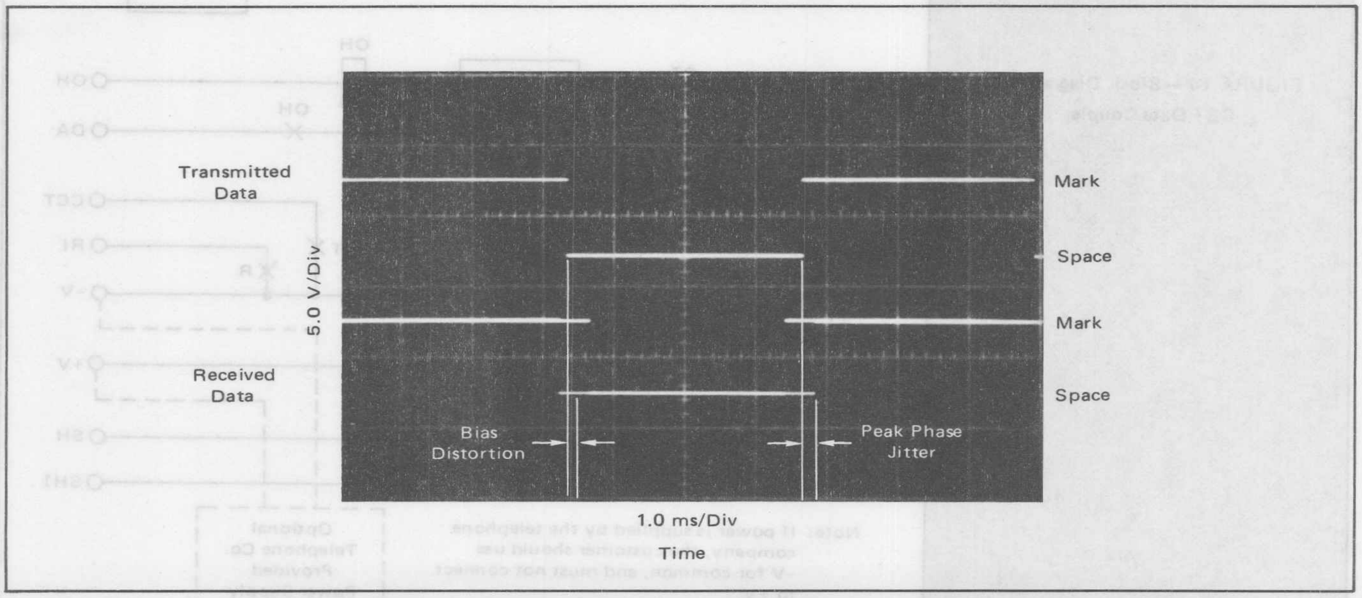

FIGURE 12 — Bias Distortion and Phase Jitter at Demodulator Output

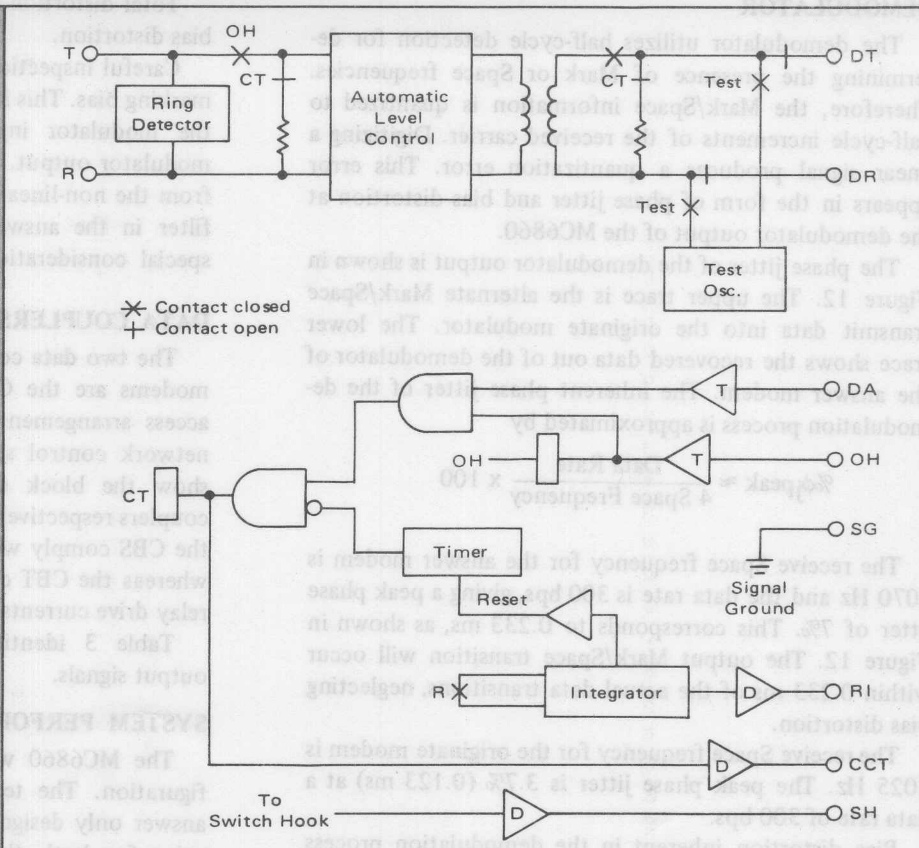

FIGURE 13 – Block Diagram of CBS Data Coupler

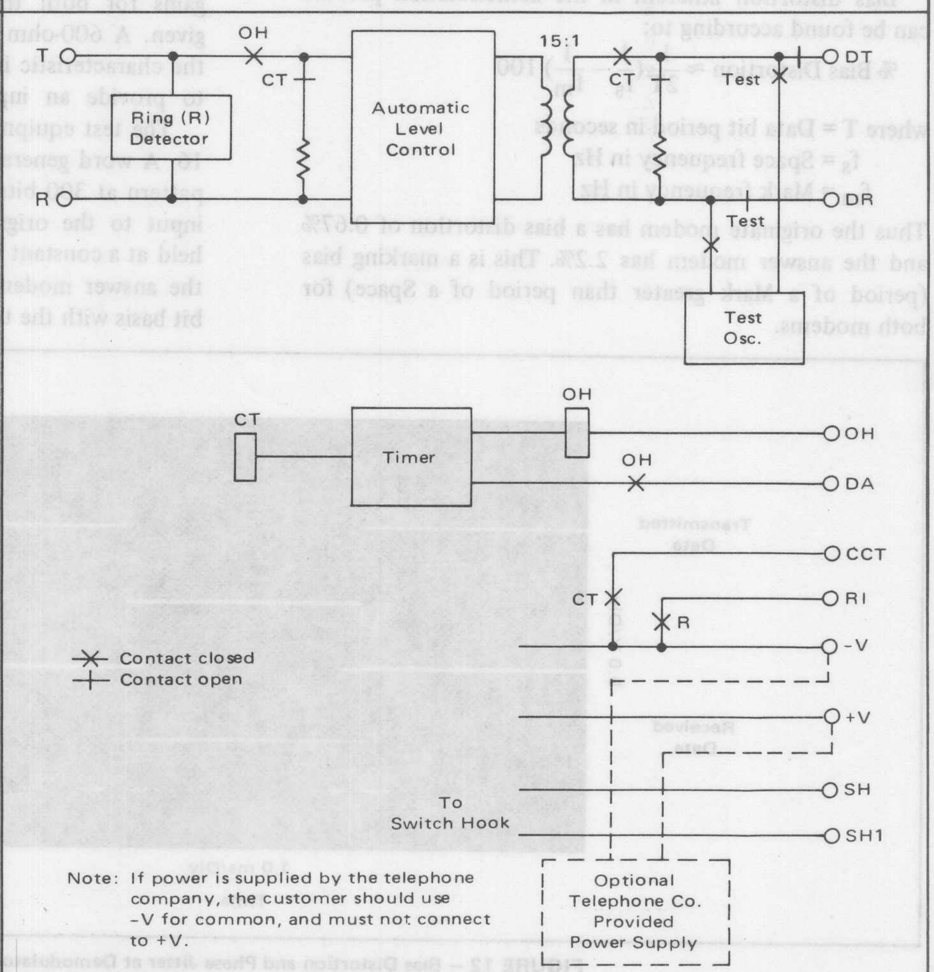

FIGURE 14 – Block Diagram of CBT Data Coupler

TABLE 3 – Data Coupler Interface Signals

Lead Designation			
Voltage (CBS)	Contact (CBT)	Direction	Function
DT DR	DT DR	Both	600-ohm transmission leads for data signals
OH	OH	To coupler	Control of OFF-HOOK relay
DA	DA	To coupler	To request data transmission path cut through
RI	RI	To customer	Ringing signal present
SG	*	Both	Signal ground in coupler (CBS)
CCT	CCT	To customer	Coupler transmission path cut through
SH	SH	To customer	Status of telephone set switch hook
*	SH1	To customer	Return for SH lead in coupler (CBT)
*	+V	To coupler	Positive dc power to coupler (CBT)
*	–V	Both	Return for dc power and common for all contact closures except the SH, SH1 pair in coupler (CBT)

*Not used in this unit.

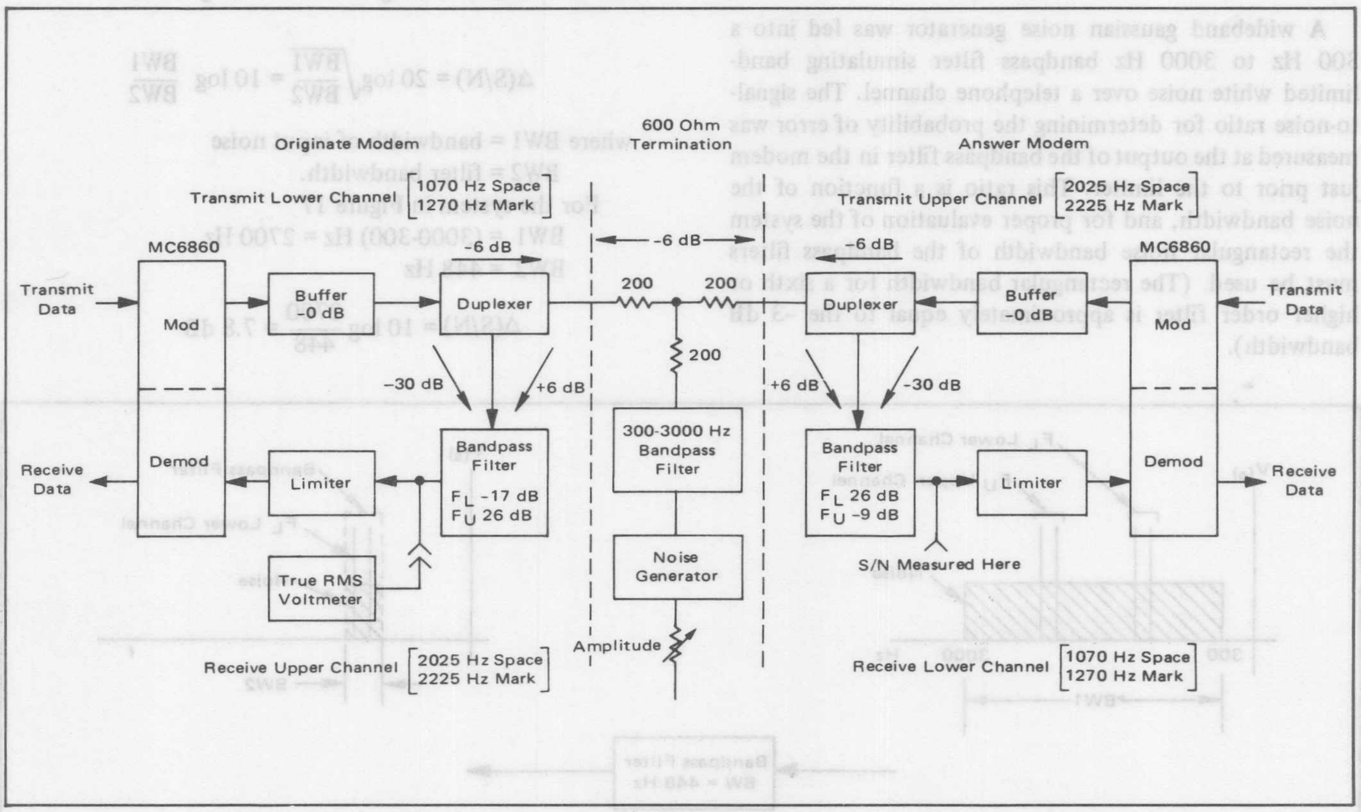

FIGURE 15 – Modem Evaluation

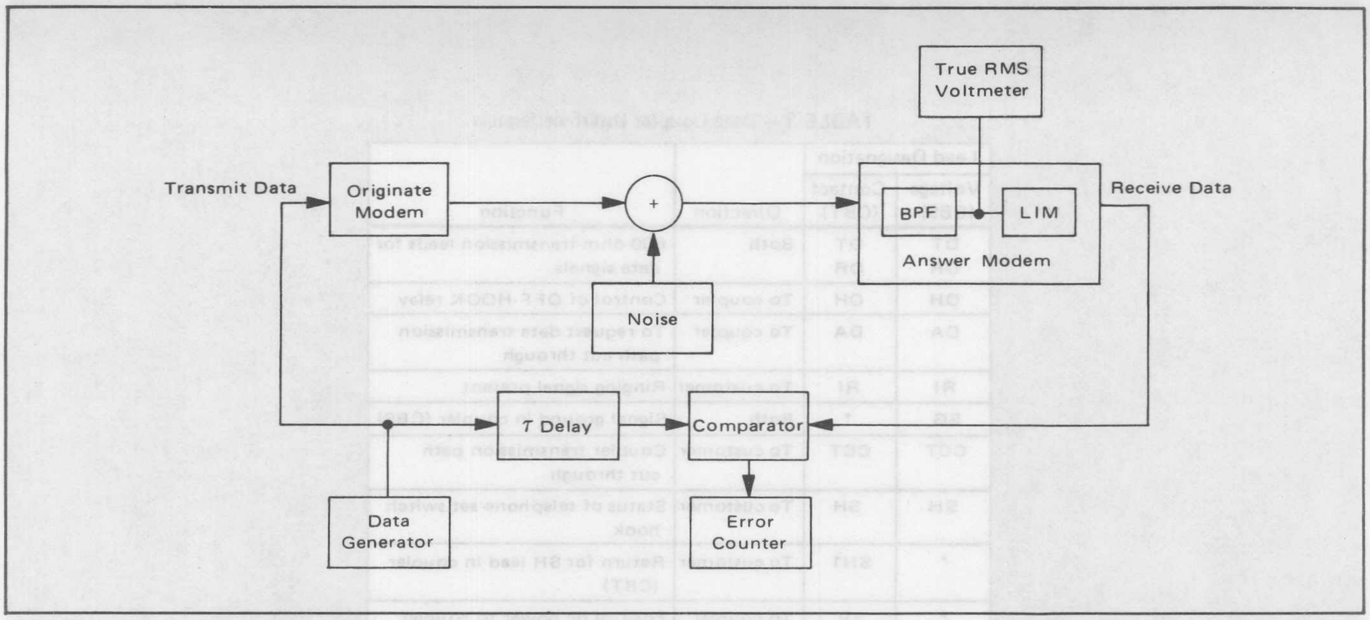

FIGURE 16 – Modem Test Equipment Configuration

Since the received data was delayed in time due to the time delay of the bandpass filter and the demodulator, the transmit data also had to be delayed an equal time before a meaningful bit-by-bit comparison could be made. This is accomplished by the τ delay between the data generator and the comparator. Sampling by the comparator was done at the center of the data bit. The number of bits used to determine the probability of error (P_e) was

$$\text{Number of bits} \geqslant \frac{100}{P_e}$$

A wideband gaussian noise generator was fed into a 300 Hz to 3000 Hz bandpass filter simulating band-limited white noise over a telephone channel. The signal-to-noise ratio for determining the probability of error was measured at the output of the bandpass filter in the modem just prior to the limiter. This ratio is a function of the noise bandwidth, and for proper evaluation of the system the rectangular noise bandwidth of the bandpass filters must be used. (The rectangular bandwidth for a sixth or higher order filter is approximately equal to the -3 dB bandwidth).

The signal and noise spectrum on the simulated transmission line is shown in Figure 17. Both the lower channel, F_L, and the upper channel, F_U, are present, along with the additive noise. When these signals are fed to the bandpass filter centered about F_L, all signals outside the passband are attenuated as shown.

The total amount of noise and F_U energy relative to the energy of F_L has now been reduced, thus improving the signal-to-noise ratio. The improvement of F_L to noise can be found according to the following formula:

$$\Delta(S/N) = 20 \log \sqrt{\frac{BW1}{BW2}} = 10 \log \frac{BW1}{BW2}$$

where BW1 = bandwidth of input noise
BW2 = filter bandwidth.

For the system in Figure 17

BW1 = (3000-300) Hz = 2700 Hz

BW2 = 448 Hz

$$\Delta(S/N) = 10 \log \frac{2700}{448} = 7.8 \text{ dB}$$

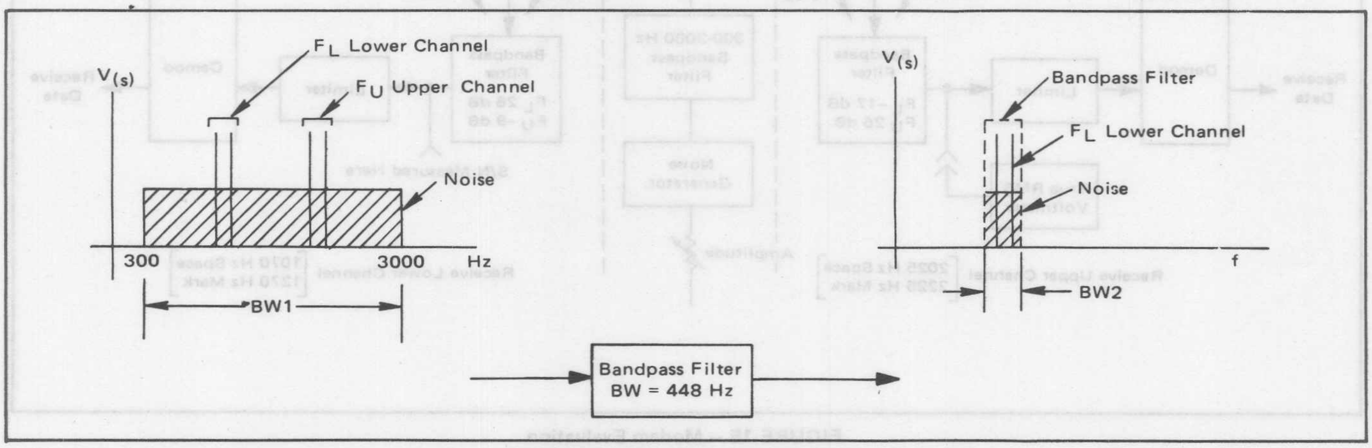

FIGURE 17 – Signal-to-Noise Improvement

Thus a signal-to-noise ratio of 12 dB at the filter output corresponds to 4.2 dB at the filter input.

$$(S/N)_{BW2} = (S/N)_{BW1} + \Delta(S/N)$$

The result of the performance tests is given in Figure 18. The theoretical probability of error (P_e)[1] curve for non-coherent FSK is determined according to;

$$P_e = \tfrac{1}{2} e^{-\frac{1}{2}\left(\frac{V_s}{V_n}\right)^2 \left(\frac{BW_n}{BW_s}\right)}$$

where V_s = signal level

V_n = noise level (true rms)

BW_n = rectangular noise bandwidth

BW_s = signal bandwidth = 1/bit time = 1/T

The dashed curve in Figure 18 is based on the signal bandwidth and rectangular noise bandwidth being equal. Since the signal bandwidth is 300 Hz (300 bits per second) and the bandpass filter of the test circuit has a measured bandwidth of 448 Hz, the theoretical P_e curve now shifts to the left by the amount of

$$\Delta(S/N) = 10 \log \frac{448}{300} = 1.74 \text{ dB}$$

The measured P_e curve deviates from the theoretical by approximately 0.5 dB. In order to maintain a $P_e \leqslant 1 \times 10^{-5}$, a signal-to-noise ratio at the limiter input must be greater than 12.2 dB. This corresponds to a signal-to-noise ratio on the telephone line of 4.4 dB in a 2700 Hz bandwidth or a signal-to-noise ratio of 3.94 dB in a 4000 Hz bandwidth.

SUMMARY

This application note describes the basic functions of a low speed FSK modem using the MC6860. The criteria for design of each function are presented. A typical test configuration is illustrated and the results are documented. The interface to standard data couplers is also included.

ACKNOWLEDGEMENT

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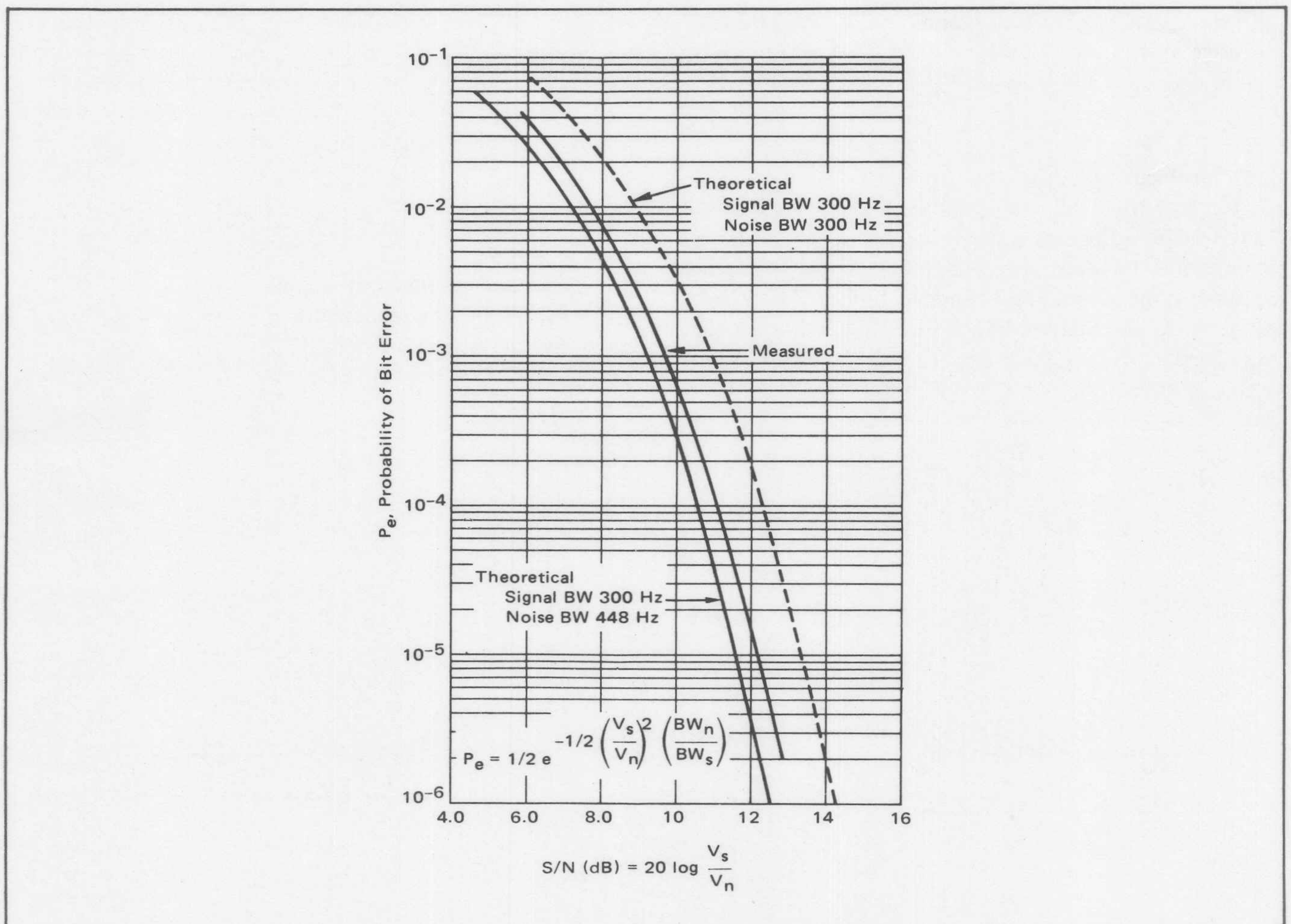

FIGURE 18 – System Performance with Gaussian Noise